APPLICATION FOR A UNITED STATES PATENT

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Title: **METHOD AND SYSTEM FOR A MULTIPLE DIMENSIONAL ADAPTIVE FREQUENCY DOMAIN NOISE CANCELER FOR DMT TRANSCEIVERS**

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CROSS REFERENCES TO RELATED APPLICATIONS

The present application is a continuation–in–part of co–pending U.S. Patent Application Serial No. 09/628,842 filed July 31, 2000.

FIELD OF THE INVENTION

5   This present invention relates to reducing the effect of narrowband noise in a multi-carrier transmission system.

BACKGROUND OF THE INVENTION

In today's modern world, businesses and residential users are demanding faster network access to the Internet. The high demand for faster network access is putting

10   pressure on vendors and service providers to choose network transmission technologies that will satisfy the emerging demand. The choice of network transmission technologies is critical since it may affect service, cost, and ultimately vendor/service provider success.

Many of the vendors and service providers have chosen to pursue digital

15   subscriber line ("DSL") technology and more specifically asymmetrical DSL ("ADSL") for providing fast Internet access to business and residential users. ADSL often provides high-speed data transmission over standard telephone lines while maintaining voice traffic on the same lines. ADSL can be seen as a cost-effective alternative to other network transmission technologies.

20   ADSL technology often exploits the relatively high bandwidth of copper loops by converting twisted-pair copper telephone wires into paths for multimedia, data communications, and Internet access. Typically, ADSL supports 1.544 to 6 Mbps

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transmission downstream and 640 kbps upstream. ADSL service may be provided by connecting a pair of modems, one often located in the telephone company's central office ("CO") and the other located at the customer premises, over a standard telephone line.

An ADSL modem, utilizing American National Standards Institute ("ANSI")

5   appointed discrete multitone ("DMT") as the modulation scheme, segments the frequency spectrum on a copper line into 256 channels. Each 4 kHz channel is capable of carrying up to 15 data bits according to the ANSI Standard T1.413, the contents of which are incorporated herein by reference. A similar standard, Recommendation G.992.1 from the International Telecommunication Union ("ITU"), is also incorporated herein by reference.

10 reference. A variation of the standard that accommodates POTS service without the use of a signal splitter is set forth in ITU Specification G.lite, or Recommendation G.992.2, the contents of which are incorporated herein by reference.

Typically, during channel analysis, a wide-band test signal sent over the 256 channels is transmitted from the ADSL terminal unit ("ATU-C") at the CO to an ADSL

15 remote terminal unit ("ATU-R") at the customer premises. The ATU-R measures and updates the noise content of each of the channels received and then determines whether a channel has sufficient quality to be used for further transmission. Depending on the quality, the ATU-R may instruct the ATU-C how much data this channel should carry relative to the other channels that are used. Often, this procedure maximizes performance

20 and minimizes error probability at any data specific rate. For instance, with a DMT modem, bit distribution may avoid noise by not loading bits onto channels that are

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corrupted by Amplitude Modulation ("AM") radio interference. The DMT modem may also lower bit distribution at the frequencies where notching occurs.

However, there are nearly 5,000 AM radio stations licensed in the U.S. to broadcast at frequencies between 540 kHz and 1.7 MHz. Unfortunately, ADSL service

5   providers typically use the frequencies between 25 kHz and 1.1 MHz to download and upload data. This sizeable overlap, approximately 560 kHz of bandwidth, can cause electromagnetic conflict because AM radio and ADSL try to use the same electromagnetic frequencies at the same time. Thus, as explained earlier, ADSL modems typically stop using the segment of the frequency spectrum occupied by any nearby AM

10   stations. Therefore, when an AM signal interferes with a carrier, a current remedy is to stop using that carrier, which consequently reduces the bandwidth and data throughput.

Additionally, the longer a wire is from the central office to the remote terminal, the more susceptible the ADSL line is to interference, especially as the signal gets weaker as it travels down the wire. Moreover, the effect is particularly pronounced if the AM

15   transmitter is near the remote terminal at the end of a long wire.

Interference caused from AM radio stations is part of a group commonly referred to as narrowband interference. Narrowband interference includes a signal whose essential spectral content may be contained within a voice channel on nominal 4-kHz bandwidth such as found in Amateur radio, AM, and Frequency Modulation ("FM") radio

20   signals.

For example, consider an AM transmission occurring at the frequency of 1070 kHz. If an ADSL signal is at the same frequency in a wire, then the ADSL receivers at

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the end of the wire may pick up the AM signal at 1070 kHz. To avoid this interference, data is often not transmitted on that particular frequency and its neighboring frequencies, because energy from the interference can also leak into signals centered on nearby frequencies. This can cause a reduction in possible throughput of the communication

5   channel. Nevertheless, this technique is currently used by the modulation standard of ADSL T1.413.

Thus, there is a need to reduce narrowband interference to increase throughput in a multi-carrier communications.

SUMMARY OF THE INVENTION

The system and method of the preferred embodiments may be directed to improving the signal-to-noise ratio in frequency spectrum regions where narrowband interference may be present. The system and method of the preferred embodiments includes reducing the narrowband interference by determining a noise estimate. In accordance with the noise estimate and output of a frequency domain equalizer, a noise-cancelled output may be obtained.

In accordance with one aspect of the present invention, a method for improving the signal-to-noise ratio in frequency spectrum regions where narrowband interference may be present includes the step of receiving at least one decoder error for the at least one carrier. The step of determining at least one adaptive, one-dimensional or two-dimensional, filter tap for each of the at least one carrier in relation to the received decoder error(s). The step of forming a noise estimate relating to the decoder error(s) and the adaptive filter tap(s). The step of receiving an FEQ output in relation with a frequency domain equalizer. Finally, the step of determining a signal having increased signal-to-noise ratio in response to the noise estimate and the FEQ output.

In accordance with another aspect of the present invention, a device for increasing a signal-to-noise ratio for at least one carrier in a multicarrier transceiver includes a canceller and a symbol storage unit. The canceller receives at least one decoder error for the at least one carrier and an FEQ output in relation with a frequency domain equalizer. The symbol storage unit stores the at least one decoder error. The canceller may then

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determine at least one, one-dimensional or two-dimensional adaptive filter tap for each of the at least one carrier in accordance with at least one stored decoder error and forms a noise estimate relating to at least one decoder error and the adaptive filter tap(s).

In a preferred embodiment, the reduction of narrowband interference is performed by a DMT receiver utilizing ADSL protocol. In another preferred embodiment, the receiver utilizes DSL protocol and any DSL variation protocol such as ADSL, very high data-rate DSL ("VDSL"), high bit-rate DSL ("HDSL"), and rate-adaptive DSL ("RADSL").

The foregoing and other objects, features and advantages of the system and method for reducing narrowband interference will be apparent from the following more particular description of preferred embodiments of the system and the method as illustrated in the accompanying drawings.

BRIEF DESCRIPTION OF THE DRAWINGS

Preferred embodiments of the present inventions are described with reference to the following drawings, wherein:

FIG. 1 is a diagram illustrating an exemplary receiver of the preferred embodiments;

FIG. 2 is a flow diagram illustrating a preferred embodiment of a method for reducing narrowband interference in accordance with the preferred embodiments;

FIG. 3 is a diagram illustrating an exemplary execution of the method in FIG. 2

FIG. 4 is a diagram illustrating exemplary receiver components in accordance with the preferred embodiments; and

FIG. 5 is diagram illustrating an exemplary use, by a two-dimensional filter, of complex taps that span the frequency domain symbol over both time and frequency.

FIG. 6 is diagram illustrating an exemplary use of the complex taps in FIG. 5 to calculate a complex tap at bin 10.

FIG. 7 is a plot of signal-to-noise ratio to bin number of a receiver in accordance with preferred embodiments.

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DETAILED DESCRIPTION OF PREFERRED EMBODIMENTS

The system and method of the preferred embodiments is directed towards improving the signal-to-noise ratio in frequency spectrum regions where narrowband interference may be present. The system and method of the preferred embodiments includes reducing the narrowband interference by determining a noise estimate. In accordance with the noise estimate and output of a frequency domain equalizer, a noise-cancelled output may be obtained.

The system and method including a discrete multi-tone receiver ("DMT receiver") have been implemented in a communication system compatible with ADSL transmission protocols, as set forth in ANSI specification T1.413. However, the receiver and method may be well suited for other multi-carrier, discrete multi-tone, or orthogonal frequency division modulation ("OFDM") systems.

In a digital transmission system preferably providing ADSL service, a DMT transceiver at a central office ("CO") is interfaced with a variety of digital services such as telephony, video-on-demand, video conferencing, and the Internet. The DMT transceiver located at the CO referred to as the ADSL transmission central office unit ("ATU-C") relays the variety of services in the form of data to a DMT transceiver located at a customer's premise such as a home or business location. The DMT transceiver at the customer's premise or remote terminal ("RT") is referred to as the ADSL transmission remote unit ("ATU-R"). The ATU-R may be connected to a computer or other application device such as a TV, audio equipment, and less intelligent devices (i.e.,

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thermostats, kitchen appliances, etc.). The ATU-C and the ATU-R typically connected together over a telephone line preferably transmit and receive data.

FIG. 1 illustrates an exemplary receiver 100 located at the RT utilizing the method and device of a preferred embodiment. The receiver 100 may be combined with a transmitter (not shown) to form a DMT transceiver or an ATU-R. It should be understood that the receiver 100 may include less or more elements such as a time domain equalizer ("TEQ"), echo canceller, and may include more or less filters.

The receiver 100 receives an analog signal $r(t)$ that has been transported over a communication channel 104 typically from an ATU-C, ATU-R, or any other DMT transceiver. The analog signal $r(t)$ may pass through a high pass filter 108 to provide frequency band separation, additional noise rejection, and/or pre-emphasis filtering. Pre-emphasis filtering preferably equalizes the frequency spectrum and may be performed by a hardware filter or with a software operation. In either case, gain may be accomplished for all frequencies in a range of pre-selected frequencies. Furthermore, the analog signal $r(t)$ may pass through an anti-aliasing, low pass filter 112 prior to sampling. It should be understood that the preferred embodiments are not to be limited by the number, or type, of filters shown in the receiver 100.

Preferably, the receiver 100 demodulates the signal $r(t)$ at a rate $\hat{f}s$. The rate $\hat{f}s$ is an estimated sampling rate utilized by the analog to digital converter ("ADC") 110 preferably matching the rate of data sent out of the digital to analog converter ("DAC")

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utilized in a transmitter. Preferably, the sampling rate reduces undesirable synchronization errors thus reducing signal attenuation and phase rotation.

The receiver 100 may then process the demodulated signal (i.e., digital samples) by converting the samples from a serial fashion into a parallel fashion and removing a

5   cyclic extension (if the cyclic extension was previously added onto the signal). Typically, the conversion of data into the parallel fashion is performed by a serial-to-parallel converter 116 ("S/P converter"). The cyclic prefix may be added at the transmitter by taking samples from an end of a data block and copying the samples to the beginning of the symbol. The cyclic prefix may then operate as a guard space between

10   neighboring transmit symbols in the time domain thus combating intersymbol interference ("ISI") efficiently in the time domain. Additionally, the periodicity of the transmitted signal, due to the cyclic prefix, enables cyclic convolution between the channel impulse response and the transmitted signal to be simulated.

Thus, the channel effect is reduced to an element-by-element multiplication

15   between the Fourier transforms of the channel impulse response and the transmitted signal, therefore introducing only different gains and delays on each carrier. These different gains and phases may be handled by a one-tap per channel equalizer (described in more detail below) thus reducing or eliminating inter-carrier interference ("ICI"). Preferably the cyclic prefix is used in the data transfer between a transmitter and the

20   receiver 100, but the preferred embodiment is not limited to utilizing the cyclic prefix. Other methods for reducing ISI and ICI, as is known in the art, may be utilized.

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The incoming serial stream of samples is converted into blocks of parallel data with N parallel values. These are fed into an N-point FFT module 120, therefore transferring the time domain signal again into the frequency domain. The transfer into the frequency domain may also mean the separation of the N/2 parallel independent carriers whose contents can now be further processed on a per bin basis. One of the N/2 outputs is commonly referred to as a bin, where the FFT module 120 outputs may then output N/2 number of bins.

A frequency-domain equalizer ("FEQ") 124 performs one-tap per channel equalization by multiplying the FFT outputs with a single complex tap. Typically, the FEQ 124 adaptively scales each subchannel by the inverse of the channel gain and phase so that a common decision boundary may be used in decoding the received data. The channel gain and channel phase typically result from the copper line between the ATU-C and the receiver 100 distorting the signal amplitude and phase, a distortion that changes from carrier to carrier. The frequency equalizer is designed to correct this channel attenuation and phase shift. The FEQ rotates the received constellation at each tone for channel phase compensation and increases the received amplitude in order to correct loop attenuation. It should be understood that the receiver 100 might utilize any type of equalizer that performs the equivalent of the FEQ.

The resulting output of the FEQ 124 may then be processed by a canceller 128. Canceller 128 may reduce noise on individual bins containing signals sent by a transmitting DMT transmitter. Canceller 128 may be activated on a bin if the correlating carrier is subjected to narrowband noise interference such as AM radio, FM radio, and/or

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any signal whose essential spectral content may be contained within a voice channel on nominal 4-kHz bandwidth. To determine if canceller 128 should be activated, the signal-to-noise ratio is preferably measured on each bin, and the bins with the smallest signal-to-noise ratio are candidates for narrowband interference cancellation. In addition, dips or nulls in the signal-to-noise distribution may be used to identify bins subjected to narrowband interference. Canceller 128 may reject and or compensate for the interference, thus desirably enhancing data throughput over the transmission channel 104. Data throughput is preferably enhanced, because a particular bin, or neighboring bins, experiencing narrowband noise interference may be utilized in data transfer, with noise cancellation, and are not deactivated due to the noise interference. It should be understood, however, that additional methods, known in the art, that may be used to determine if a bin or group of bins are experiencing narrowband interference.

Symbol storage 140 may be utilized to store decoder 132 outputs including canceled outputs, where canceled outputs are direct outputs from the decoder 132, or uncanceled decoder outputs, where uncanceled outputs are outputs from the decoder 132 with an added noise estimate. Symbol storage 140 may include RAM, hard disk, EEPROM, ROM, etc. If noise cancellation is necessary, the canceller 128 may utilize information from the decoder 132 and the outputs of the FEQ 124. Additional processing may be performed on the output of the decoder 132, such as Reed Solomon coding and ATM decoding 136. It should be understood that the receiver 100 is not limited to that shown in FIG. 1, it may include more or less elements such as additional filters, a means for echo cancellation, a time domain equalizer, and so on.

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FIG. 2 illustrates a top-level flow diagram of an embodiment for rejecting, or compensating, or both for narrowband interference. The system and method shown in FIG. 2 may be applied to a canceller 128, including a symbol storage unit 140. The system and method may be applied in the form of executable software read by the

5 canceller 128 from a memory device, such as ROM, RAM, EEPROM, hard disk, etc. Additionally, the system and method shown in FIG. 2 may be applied in the form of active, passive, and/or logic devices such as comparators, shift registers, adders, etc.

The system and method for reducing narrowband interference in accordance with the preferred embodiments includes receiving and storing at least one decoder error

10 originating from a decoder (see, for example, 132 in FIG.1). In FIG. 3, the decoder error including a canceled decoder error or an uncanceled decoder error may be either stored or equivalently delayed for a predetermined amount of time (FIG. 3 shows delayed uncanceled decoder errors). The decoder errors may be stored in the symbol storage (see, for example, 140 in FIG. 1) as represented by the delays 300,304.

15 The number of decoder errors stored may be related to the desired size of the adaptive filter. For example, one decoder error might be stored and utilized for a one-tap filter; J decoder errors might be stored and utilized for a J-tap filter. The decoder error(s) may include slicer errors such as caused by a phase shift in a constellation of symbols. It should be noted that the canceled error may be utilized instead of the uncanceled error,

20 but convergence of the adaptive filter taps may be slower.

The method further includes the step 204 of determining adaptive filter taps. The adaptive filter taps 308,312, such as shown in FIG. 3, may be initially calculated during

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the MEDLEY phase of initialization to minimize either the sum of squared errors over M

symbol periods and/or, the mean squared error ("MSE"). The MEDLEY stage includes

estimation at the ATU-R of the downstream signal-to-noise ratio ("SNR"), that is, the

SNR of the signal from the ATU-C to the ATU-R.

5   The filter taps may be continuously adjusted during receiver 100 operation. The

least means square ("LMS") may be utilized to update the filter taps which can minimize

the MSE given by the relation: $\xi = E\left\{ \left| f(n)y(i)\text{-}x(i) - \bar{h}_k \bar{e}(i)^T \right|^2 \right\}$ where $x(i)$ is a known

transmitted symbol, such as during receiver training and/or the decision for the current

constellation during showtime, or equivalently, the steady state signaling state, where $f(n)$

10   is the FEQ coefficient corresponding to the Nth bin, where $y(i)$ is the FFT output of the

corresponding Nth bin, and where $\bar{h}_k \bar{e}(i)^T$ is the finite impulse response ("FIR") or

infinite impulse response ("IIR") filtering of the uncanceled decoder error row vector

$\bar{e}(i)$ with filter coefficient row vector, $\bar{h}_k$.

The uncanceled decoder error row vector given above $\bar{e}(i)$ (in the above MSE

15   relation) is written in the transpose form, signifying that if $\bar{h}_k$ is in a row vector form,

then $\bar{e}(i)$ should be in column vector form. Of course, it is possible to have both vectors

$\bar{h}_k$ and $\bar{e}(i)$ in column vector form, in which, the impulse response could be written as

$\bar{h}_k{}^T \bar{e}(i)$ such that the desirable result of a (row vector) * (column vector) = (scalar result)

is achieved. Therefore, it should be understood that by showing $\bar{e}(i)$ in the transpose

20   form, as shown above and below, does not limit $\bar{e}(i)$ to the transpose form, but simply

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designates that $\bar{\hat{e}}(i)$ and $\bar{h}_k$ are not alike in form, unless it is specifically specified as such; and that when the two vectors are multiplied together, they can preferably form a scalar value.

In an exemplary embodiment, coefficient vector, $\bar{h}_k$, may include one coefficient

5   (shown as H(0) 308 in FIG. 3) or up to J coefficients (shown as H(J) 312 in FIG. 3) and may be given by the row vector: $\bar{h}_k = [h_0, h_1, ..., h_{J-1}]$ where k is an index counter. Uncanceled decoder error vector, $\bar{\hat{e}}(i)$, is the uncanceled decoder error for the symbol i of a total J symbols and may be given by the transpose of the row vector: $\hat{e}(i) = [\hat{e}(i-1), \hat{e}(i-2), ..., \hat{e}(i-J)]$.

10   The uncanceled decoder error may be found in part from the canceled decoder error vector $\bar{e}(i)$ and from the noise estimate vector $\bar{n}_r(i)$. It should be noted that the canceled decoder error vector may be utilized instead of the uncanceled decoder error vector, but convergence of the adaptive filter taps may be slower. To use the canceled decoder error vector, the canceled decoder error vector $\bar{e}(i)$ may be substituted for the

15   uncanceled decoder error $\bar{\hat{e}}(i)$ in any of the relationships described herein.

To minimize the MSE, the adaptive filter taps may be determined in accordance with the relation: $\bar{h}_{k+1} = \bar{h}_k + \alpha(f(n)y(i) - x(i) - \bar{h}_k\bar{\hat{e}}(i)^T)conj(\bar{\hat{e}}(i))$, where $x(i)$ is a known transmitted symbol, such as during receiver training and/or the decoder decision for the current constellation during showtime, where $f(n)$ is the FEQ coefficient corresponding to

20   the Nth bin, where $y(i)$ is the FFT output of the corresponding Nth bin, where $\bar{h}_k$ is the

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coefficient vector, where α is the corrective coefficient, where $\bar{h}_k \bar{e}(i)^T$ is the filtering of

the constellation error vector $\bar{e}(i)$ with filter coefficient vector $\bar{h}_k$, and where $conj(\bar{e}(i))$

is the complex conjugate of the input signal to the filter which appears in the LMS

adaptive update term for the symbol i for a total J symbols.

5   In the exemplary embodiment, the corrective coefficient may be calculated during

the R_REVERB3 stage of receiver initialization. (R_REVERB3, as is known in the art, is

a latter stage of receiver initialization typically used to measure the upstream power,

adjust receiver gain control, synchronize the receiver, and train the FEQ). Additionally,

it may be possible to determine the corrective coefficient concurrently with the training

10   of the FEQ (see, for example, 124 in FIG. 1).

In another embodiment, a single tap predictor may be utilized to determine a filter

tap and is found in accordance with the relation: $h_{k+1} = \alpha h_k + (1-\alpha)\dfrac{\hat{e}(i)}{\hat{e}(i-1)}$ where α is

the corrective coefficient, where h_k is the adaptive filter tap for symbol i, and where

$\hat{e}(i)$ is the uncanceled decoder error for the symbol i. The single tap predictor provides

15   an estimate of the current interface component by rotating and scaling the previous slicer

error. The single tap predictor may be updated with past rate-of-change information (that

is, the uncanceled decoder error rate of change) in an attempt to whiten the current slicer

error.

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The uncanceled decoder error(s) is filtered per step 208 to create a noise estimate per step 212. The noise estimate (shown because of filtering 316 the uncanceled decoder error(s) in FIG. 3) may be determined in accordance with the relationship:

$$\bar{n}_R(i) = h_0\hat{e}(i-1) + h_1\hat{e}(i-2) + ... + h_{J-1}\hat{e}(i-J) = \bar{h}_k\bar{e}(i)^T$$

where $\bar{e}(i) = [\hat{e}(i-1), \hat{e}(i-2), ..., \hat{e}(i-J)]$, and where $\bar{h}_k = [h_0, h_1, ..., h_{J-1}]$. The uncanceled decoder error(s) may be filtered using any available filtering technique such as FIR or IIR filters.

Further, a received FEQ output of the Nth bin per step 216 is combined with the noise estimate for the corresponding Nth bin to create a canceller output 220 for the Nth bin preferably with an increased signal-to-noise ratio.

It should be understood that the exemplary flow diagram provided in FIG. 2 and the exemplary block diagram in FIG. 3 are not limited to the steps shown and that other steps while remaining within the scope of the invention may be utilized. Furthermore, the steps may not have to be performed in the order as shown in FIGS. 2 and 3.

In a preferred embodiment, shown in FIG. 4, a canceller 128 improves the signal-to-noise ratio by reducing or eliminating the narrowband noise interference on a per bin basis. The output of the canceller 128 can be further processed by a decoder 132 such as by a slicer. The decoder 132 preferably provides the decoder errors, which are stored in a storage device, referred to as a symbol storage unit 140. The symbol storage 140 unit may include any device in which the decoder error(s) may be stored in such as, for example, but not limited to, a random access memory ("RAM"), a buffer, and an

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electrically erasable programmable read-only memory ("EEPROM"). The canceller 128 may then utilize the stored decoder errors and the output 144 of the FEQ to reduce narrowband interference on a per bin basis.

Assume, for example, bin L (shown in FIG. 4) has a low signal-to-noise ratio. To increase the signal-to-noise ratio, the canceller 128 has determined to reduce the noise, thus increasing the signal-to-noise ratio of bin L. To reduce the noise of bin L, output from the decoder 132 is filtered and added with a noise estimate and stored in the symbol storage device 140. The data stored in the symbol storage 140 may include up to J decoder errors originating from the decoder 132 taken from bin L. The decoder errors or symbols stored in the symbol storage 140 may be used to develop adaptive filter taps to reduce undesired interference. This can be accomplished by filtering the decoder errors with an adaptive filter having up to J taps. Once filtered, the interference on bin L is preferably reduced. The noise cancelled data from bin L may be used in further processing such as a Reed Solomon/ ATM decoder (not shown).

As described above, narrow band interference in the passband of communicating DMT modems can be modeled as additive noise. The interference can usually be correlated and predicted, which suggests that in the frequency domain, noise interference is correlated within each symbol in time and with adjacent carriers within the current symbol. Thus, in another embodiment, a two-dimensional ("2-D") adaptive FIR filter with complex taps spanning the frequency domain symbol over both time and frequency can be used to predict the additive noise component for each carrier. In this embodiment,

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a 2-D FIR filter is a 2-D tapped delay line defined over an arbitrary 2-D filter mask, which can be square, rectangular, or any other shape.

Referring to FIG. 5 is a diagram illustrating an exemplary tapped delay line for n/2 bins outputted from the FEQ 124. Preferably, delayed and current taps for symbols of the outputted FEQ 124 bins are utilized to define the 2-D filter. Thus, pointers i and j are used to designate a decoder error corresponding to a related symbol and bin, respectively, of interest. Therefore, according to this example, for bin 0, or $j = 0$, the tap for symbol block 0, or $i = 0$, is the tap calculated for the most recent symbol in time, whereas the tap in block 6, or $i = 5$, is the tap calculated for the older symbol in time. This process of storing or delaying decoder errors can occur for each bin and symbol, in which according to this illustration, up to $(N/2)(6)$ decoder errors would be stored or delayed.

As more new symbols are outputted from the FEQ 124, the decoder errors that were once in the blocks labeled 0-5 are consequently shifted right, such that the most recent decoder error is placed in block 0. More or fewer decoder errors can be utilized, but have been limited to 6 (labeled 0-5) in this illustration for purposes of clarity and ease of demonstration.

The 2-D filter taps may be continuously adjusted during receiver operation. Accordingly, the LMS may be utilized to update the 2-D filter taps which can minimize the MSE and is given by the following relation:

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$$\xi = E\{e(i,j)^2\} = E\{(f(i,j)y(i,j)\text{-}x(i,j) - \bar{h}_{ij}\,\bar{\bar{e}}(i,j)^T)^2\}$$ where $x(i,j)$ is a receiver known copy of

the transmitted data for the jth bin and the ith symbol, such as determined during receiver

training and/or the decision for the current constellation during showtime, where j

corresponds to a bin and the i corresponds to a symbol, $f(i,j)$ is the FEQ coefficient, $y(i,j)$

is the FFT output, and $\bar{h}_{i,j}\,\bar{\bar{e}}(i,j)^T$ is the FIR or IIR filtering of the uncanceled decoder error

vector $\bar{\bar{e}}(i,j)$ with filter coefficient vector $\bar{h}_{i,j}$.

Although the 2-D filter can utilize a filter mask of any shape including a square or

rectangular, the 2D filter coefficients $h_{i,j}$ for a square filter mask may be given by the

following row concatenated coefficient vector $\bar{h}_{i,j}$:

$$\bar{h}_{i,j} = \left[h(0,-N_f)...h(N_t,-N_f)...h(1,0)...h(N_t,0)...h(0,N_f)...h(N_t,N_f)\right]$$

where $h_{i,j}(k,l)$ is the filter tap for the ith symbol of the jth bin, displaced by k symbols and l

bins.

Preferably, each bin that has an active noise canceller can have its own set of 2-D

coefficients. The filter mask, $\bar{h}_{i,j}$, defined above is rectangular in shape (more of which

is described below) and typically symmetric about the bin designated as the jth bin and

contains $2N_f+1$ rows and N_t+1 columns for a total of up to $(2N_f+1)(N_t+1)-1$ taps, where

N_f designates the number of bins spanned on each side of the target bin being cancelled,

and where $N_t + 1$ designates the number of taps in the symbol delay line for each bin,

except for the target bin that has N_t taps.

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Figure 6 is a diagram illustrating an exemplary embodiment for calculating $\bar{h}_{i,j}$ for bin 10. It should be understood, however, that the present embodiments can calculate $\bar{h}_{i,j}$ for any bin, and that in this example, bin 10 was chosen for purposes of illustration only. For this particular example, $j = 10$ and $i = 0$, therefore the filter tap for $\bar{h}_{0,10}(0,0)$ is calculated. Also, for this example, assume that $N_f = N_t = 2$. Thus, below is an exemplary coefficient vector, $\bar{h}_{0,10}(0,0)$, and is shown as the transpose of a column vector for purposes of readability:

$$\bar{h}_{0,10}(0,0) = \begin{bmatrix} h(0,-2) \\ h(1,-2) \\ h(2,-2) \\ h(0,-1) \\ h(1,-1) \\ h(2,-1) \\ h(1,0) \\ h(2,0) \\ h(0,1) \\ h(1,1) \\ h(2,1) \\ h(0,2) \\ h(1,2) \\ h(2,2) \end{bmatrix}^T$$

The shaded blocks correspond to the taps utilized to determine $\bar{h}_{0,10}(0,0)$. Notice how the shape of this filter mask forms a rectangular shape (the shaded portion). This process can be repeated until all of the desired filter taps have been calculated.

Uncanceled decoder error row vector $\bar{e}(i,j)$ is the uncanceled decoder error for the symbol i and bin j, and may be given by the following row vector:

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$$\bar{\hat{e}}(i,j) = \left[\hat{e}(i, j - N_f)...\hat{e}(i - N_t, j - N_f)...\hat{e}(i - 1, j)...\hat{e}(i - N_t, j)...\hat{e}(i, j + N_f)...\hat{e}(i - N_t, j + N_f) \right]$$

The Uncanceled decoder error vector may be found in part from the canceled decoder error $\bar{e}(i,j)$ and from the noise estimate $\bar{n}_r(i,j)$. Similar to the 1-D filter, the canceled error may be utilized instead of the uncanceled error, but convergence of the adaptive filter taps may be slower. To use the canceled error, the canceled decoder error $\bar{e}(i,j)$ may be substituted for the uncanceled decoder error $\bar{\hat{e}}(i,j)$ in any of the relationships described herein.

To minimize the MSE, the adaptive filter taps may be determined in accordance with the relation: $\bar{h}_{i+1,j} = \bar{h}_{ij} + \alpha \left(f(i,j)y(i,j) - x(i,j) - \bar{h}_{ij}\bar{\hat{e}}(i,j)^T \right) conj\left(\bar{\hat{e}}(i,j) \right)$ where $x(i,j)$ is a known transmitted symbol, such as during receiver training and/or the decoder decision for the current constellation during showtime, $f(i,j)$ is the FEQ coefficient corresponding to the ith symbol and j^{th} bin, where $y(i,j)$ is the FFT output of the corresponding i^{th} symbol and the jth bin, where $\bar{h}_{i,j}$ is the filter coefficient vector, where α is the corrective coefficient, where $\bar{h}_{i,j}\bar{\hat{e}}(i,j)^T$ is the filtering of the constellation error vector $\bar{\hat{e}}(i,j)$ with filter coefficient vector $\bar{h}_{i,j}$, and where $conj(\bar{\hat{e}}(i,j))$ is the complex conjugate of the input signal to the filter which appears in the LMS adaptive update term for the symbol i and bin j.

The corrective coefficient may be calculated during the R_REVERB3 stage of receiver initialization. Additionally, it may be possible to determine the corrective coefficient concurrently with the training of the FEQ (124 in FIG. 1).

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The uncanceled decoder error(s) is filtered per step to create a noise estimate per step. The noise estimate (shown because of filtering the uncanceled decoder error(s) in FIG. 3) may be determined in accordance with the relationship:

$$\bar{n}(i,j) = \bar{h}_{i,j}\bar{\hat{e}}(i,j)^T = \sum_{k=0}^{N_t} \sum_{\substack{l=-N_f \\ l*k=0}}^{N_f} h_{ij}(k,l)\hat{e}(i-k,j-l)$$

5   where $\bar{\hat{e}}(i,j)$ is defined above, and where $\bar{h}_{i,j}$, is defined above. The uncanceled decoder error(s) may be filtered use any available filtering technique such as FIR or IIR filters.

The noise cancelled constellation error can be given with the following relationship:

$$e(i,j) = \hat{e}(i,j) - n(i,j) = f(i,j)y(i,j) - x(i,j) - n(i,j)$$

10   where $f(i,j)$ is the FEQ coefficient corresponding to the ith symbol of the jth bin, $x(i,j)$ is the local copy of the transmitted data (or the slicer decision), and $y(i,j)$ is the received data point.

The 2-D filter in utilities previous slicer errors within a given subchannel and several adjacent subchannels to predict the current slicer error in the targeted subchannel

15   for the current DMT symbol.

Further, a received FEQ output of the nth bin per step is combined with the noise estimate for the corresponding nth bin to create a canceller output for the nth bin preferably with an increased signal-to-noise ratio.

FIG. 7 is a simulation output in accordance with the preferred embodiments.

20   Speech data (i.e., a source of noise in this example) was modulated onto a carrier wave at

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560 kHz, which for this example is equivalent to bin 130. After receiving the carriers on the bins between approximately 50 and 250 the noise added to bin 130 (i.e., approximately 560 kHz) was so great that the signal-to-noise ratio was near zero.

Additionally, the signal-to-noise ratio exhibited a large "roll off" reducing the
5   signal-to-noise ratio of surrounding bins. Such a bin may be deactivated in ordinary DMT communications. However, utilizing the system and methods described herein, the output of the canceller increased the signal-to-noise ratio for bin 130 and the surrounding bins. By increasing the signal-to-noise ratio of bin 130, the receiver may use bin 130 to transfer data to and from a transmitter.

10   The system and method of the preferred embodiments is directed to improving the signal-to-noise ratio in frequency spectrum regions where narrowband interference may be present. The system and method of the preferred embodiments includes reducing the narrowband interference by determining a noise estimate. In accordance with the noise estimate and output of a frequency domain equalizer or equivalent, a noise-cancelled
15   output may be obtained.

It should be understood that in each equation, the vectors may take any form including row or column vectors. For example, if the uncanceled decoder error $\bar{e}(i)$ is written in a row vector form, then the filter coefficient vector $\bar{h}_k$ would be in a column vector form, so long as the result of the vector multiplication results in a scalar value,
20   unless otherwise specified. Therefore, the equations described herein are not to be limited to the form of the vectors.

It should also be understood that the programs, processes, methods and systems described herein are not related or limited to any particular type of receivers or network system (hardware or software), unless indicated otherwise. Various types of general purpose or specialized systems may be used with or perform operations in accordance with the teachings described herein.

In view of the wide variety of embodiments to which the principles of the present invention can be applied, it should be understood that the illustrated embodiments are exemplary only, and should not be taken as limiting the scope of the present invention. For example, the steps of the flow diagrams may be taken in sequences other than those described, and more or fewer elements may be used in the block diagrams. While various elements of the preferred embodiments have been described as being implemented in software, in other embodiments in hardware or firmware implementations may alternatively be used, and vice-versa.

It will be apparent to those of ordinary skill in the art that methods involved in the system and method reducing narrowband interference may be embodied in a computer program product that includes a computer usable medium. For example, such as, a computer usable medium can include a readable memory device, such as a hard drive device, CD-ROM, a DVD-ROM, or a computer diskette, having computer readable program code segments stored thereon. The computer readable medium can also include a communications or transmission medium, such as, a bus or a communication link, either optical, wired or wireless having program code segments carried thereon as digital or analog data signals.

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The claims should not be read as limited to the described order or elements unless stated to that effect. Therefore, all embodiments that come within the scope and spirit of the following claims and equivalents thereto are claimed as the invention.